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DISSEMINATION LEVEL		
PU	Public	X
PCA	Public with confidential annex	
CO	Confidential, only for members of the consortium (including Commission Services)	

EXECUTIVE SUMMARY

The main objective of the SACRA work package 2 (WP2) is to develop sensing and access techniques and algorithms for cognitive energy efficient radio.

SACRA technical annex specifies the specific aim of Task 2.3 of WP2 as MIMO algorithms which has been further subdivided into “Space-Time-Frequency Polarization Codes” and “MIMO Multiplexing Techniques”. The former was presented in D2.3 whereas the later subdivision “MIMO Multiplexing Techniques” is the focus of this deliverable D2.4.

In this deliverable, various MIMO transmission and reception strategies are proposed for cognitive radio systems. In the specific setup of MIMO cognitive radios, intelligent receiver designs are considered. In this vein of cognitive radios where interference is almost ubiquitous, in particular for cognitive radios as primary systems would not care about the presence of unlicensed devices, a novel practical interference cancellation receiver design is considered which brings very attractive gains at the cost of reasonable complexity. The gains are verified through simulation results presented in the later sections. Such techniques are primordial for primary-secondary co-existence.

CONTENTS

1 INTRODUCTION

1.1 GENERAL CONTEXT

The idea of unlicensed cognitive radios has been gaining momentum at a rapid pace in the past few years due to the availability of limited spectrum for communications and huge growth in available wireless applications and services. Thus the demand for data rates on the air is increasing exponentially and this demand is facing bottleneck in terms of available spectrum. This has caused high interest of communication engineers and researchers in making unlicensed systems practical and co-exist with present licensed systems with least disturbance in their operation.

The simultaneous utilisation of one spectrum band by the primary system (the spectrum owner or license holder) and the secondary/cognitive system (the unlicensed spectrum user) asks for new sensing and access techniques. These signal processing techniques need to be tailored for the specific mode of operation of the cognitive systems. These systems are meant to use the unlicensed spectrum in various ways, although all require that primary system does not get a real hit in its quality of service. In this vein, three regimes of operations of cognitive systems have been identified. [Goldsmith2009] has classified these regimes as underlay, interweave and overlay.

Interweave Paradigm: The ‘interweave’ paradigm is based on the idea of opportunistic communication, and was the original motivation for cognitive radio by Joseph Mitola. The idea came about after studies conducted by the FCC showing that a major part of the spectrum is not utilized most of the time. In other words, there exist temporary space-time-frequency voids, referred to as spectrum holes, that are not in constant use in both the licensed and unlicensed bands. These gaps change with time and geographic location, and can be exploited by cognitive users for their communication. Thus, the utilization of spectrum is improved by opportunistic frequency reuse over the spectrum holes. The interweave technique requires knowledge of the activity information of the noncognitive (licensed or unlicensed) users in the spectrum. The deliverables D2.1 and D2.2 have presented a wide range of sensing techniques which can be exploited to identify these time-frequency holes unused by primary systems where cognitive systems can transmit their data.

Underlay Paradigm: The underlay paradigm encompasses techniques that allow communication by the cognitive radio assuming it has knowledge of the interference caused by its transmitter to the receivers of all noncognitive users. In this setting the cognitive radio is often called a secondary user which cannot significantly interfere with the communication of existing (typically licensed) users, who are referred to as primary users. The cognitive radios in such systems are transmitting simultaneously with primary systems and hence interference is present. The main objective in this paradigm is to gain sufficient knowledge of the environment, precisely the interference which a cognitive transmitter will produce at primary receivers and keep it at reasonably low levels. This interference threshold power is sometimes called interference temperature.

Overlay Paradigm: The enabling premise for overlay systems is that the cognitive transmitters will help the primary system in conveying their own information. This requires that cognitive transmitters need to obtain the knowledge of the noncognitive users’ messages. This is certainly tricky but various scenarios and system settings have been identified where this becomes practical. The codebook information for the primary systems could be obtained, for example, if the primary users follow a uniform standard for communication based on a publicized codebook which is mostly the case for all practical systems. The information transmitted by primary users might be obtained by decoding the message at the cognitive terminals. In particular it has been observed that there are a lot of re-transmissions on the air and if cognitive terminals are able to decode the information message in the first transmission, for the later transmissions they have non-causal knowledge of the primary information. This cognitive regime of operation is the most challenging of the three identified regimes of operation for cognitive radios.

For interweave cognitive communication, the sensing techniques are primordial. Once the time-frequency holes are identified, they can be used for the transmission of information among the cognitive terminals using the standard flavour of transmission and reception algorithms. On the other hand, in underlay and overlay regimes of cognitive operation, cognitive users are transmitting/receiving data on the same time-frequency slots as the primary license-holder users and hence both systems face increased amount of interference from each other. Thus it’s the responsibility of cognitive users to adapt their transmissions in such a manner to least disturb the primary system (underlay) or even help the primary information transmission (overlay). This requires sophisticated transmission and reception algorithms at the cognitive terminals which need to be optimized under certain co-existence constraints. The interference situation at the cognitive receivers is worsened by the fact that they are facing interference from both systems and in particular primary terminals will not make any effort to alleviate the interference they produce at the cognitive users. This situation could further degrade if there are multiple secondary/cognitive systems in operation in a particular geographic location.

In its current version, this deliverable presents two major contributions. In the first contribution, an overlay cognitive radio system is considered. In this system, cognitive terminals are MIMO enabled, i.e., equipped with multiple antennas. For the objective of throughput maximization, transmission and reception strategies are considered. Rate regions are derived and plotted in this section. Rate regions are very important to understand various points of operation of a multi-user communication system. It is shown that with simple linear receivers, attractive rates are achievable for cognitive users.

In the second contribution, a bigger system setup is considered which allows the presence of multiple interferers in the geographic location of interest where the cognitive system is operating. This cognitive system is receiving interference from multiple sources which is highly likely for interweave and overlay regimes of cognitive operation. Full-blown maximum-likelihood style of signal processing becomes too prohibitive in such a situation. To cope with this, a dual stage receiver structure is proposed which in the first linear stage cancels the “weak” set of interferers and in the second, non-linear stage treats the “stronger” set of interferers. This scheme brings very attractive gains at a reasonable complexity.

For SACRA use cases, it is intended to use TVWS band as the unlicensed band. Due to lower frequencies, achieving spatial diversity becomes an issue here in terms of antenna size and of antenna separation since an antenna separation distance of half a wavelength is necessary to avoid signal correlation and electromagnetic antenna coupling. This causes the design of huge antennas for such bands. This brings about the interest of harnessing the spatial diversity through polarization. So instead of true spatial diversity, two different polarizations are captured. We have not given separate study of proposed techniques for polarization but the proposed techniques and algorithms apply verbatim to dual polarized antennas.

The rest of this deliverable is organized as follows. Section II presents the design of transmission reception strategy for MIMO cognitive radio systems in overlay regime. Section III provides a novel interference cancellation framework. This framework is fairly general and can be applied to any terminal receiving interference from multiple sources and hence becomes very attractive for cognitive receivers. Simulation results verifying the performance are presented along with both sections. Finally, Section IV gives the concluding remarks for this deliverable.

1.2 PURPOSE OF THE DOCUMENT

This document is the deliverable D2.4 of WP2.3.2 of the FP7 SACRA project. This sub work package is entitled “MIMO multiplexing techniques" and this deliverable is entitled " Design of MIMO Cognitive Transmission and Reception Techniques". It considers the studies on design and analysis of transmission and reception strategies for cognitive radios having multiple antennas. It also considers the interference mitigation techniques for such cognitive radio systems.

The goal of this document is to disseminate the proposed MIMO cognitive transmission strategies and interference cancellation algorithms in a cognitive radio framework and to demonstrate their performance. The performance of the proposed techniques is shown through simulations. This document becomes an essential contribution to WP2.3 in terms of transmission and access techniques. The interference cancellation technique in this deliverable is fairly general and highly suited to the needs of environments where the receivers would be facing interferences from multiple sources. This algorithm brings very attractive gains over normal receivers with reasonable amount of extra processing and hence could be a good candidate for future wireless receivers.

In a nutshell, the proposed transmission techniques and the receive algorithm show their benefit in terms of increased spectral efficiency for SACRA cognitive radios which is the main objective of this project. Secondly interference cancellation algorithm is a very practical cost-effective one meeting the energy efficient promise of SACRA radios.

2 MIMO TRANSMIT-RECEIVER ALGORITHMS FOR OVERLAY COGNITIVE RADIO

2.1 MAIN THEME

The two user Gaussian fading Cognitive Radio (CR) interference channel is the focus of this study which comprises of two transmitter-receiver pairs. The primary transmitter (PTx) and primary receiver (PRx) are equipped with single antenna each whereas the cognitive transmitter (CTx) and the cognitive receiver (CRx) may be equipped with multiple antennas in our setting although this pair is also restricted to single stream transmission. The channel state information is partial such that each transmitter knows its channel to the PRx but has no information about its channel to the CRx. This section proposes two simple transmission strategies focusing the CR channels in the so-called “overlay paradigm” where the CTx not only transmits its own message but helps as well transmitting the primary message. In the first strategy, the primary message is independently encoded at the two transmitters whereas the same primary message encoding is used in the second strategy which not only allows the coherent signal combining at the PRx but also the possibility of complete interference cancellation at the CRx even in the limiting case when the CRx is equipped only with two antennas. The simulation results demonstrate that the proposed strategy with same primary message encoding shows considerable performance benefit over the strategy where the primary message is independently encoded at the two transmitters.

2.2 INTRODUCTION

2.2.1 The origin and background of Cognitive Radio

The simplest instance of interference channel (IFC), where two transmitter-receiver pairs are communicating over the same communication resource without any cooperation, has been the focus of research [Carleial1978] [Sato1977] for the past many decades. The complete characterization of the capacity region of this simplest version of IFC is still an open problem although is known for some special cases, e.g., very high or very low interference regimes [Carleial1975] and [Han1981]. The best known achievable region for this channel is by [Han1981]. Very recently, IFC was studied through a new perspective bounding the capacity region to within one bit [Etkin2008].

The idea of cognitive radio (CR) originated from Mitola’s work [Mitola2000] where the CRs were intelligent communication devices sensing the environment and having the capability to co-exist with primary license-holders. Initially the mainstream CRs were considered to be the spectrum sensing devices which were able to sense and transmit over the spectrum holes. The authors in [Devroye2006] proposed “the overlay paradigm” for the cognitive radios where the cognitive transmitter (CTx) knows the message of the primary transmitter (PTx), basically destined for the primary receiver (PRx). The discussion about the scenarios under which the primary message can be estimated at the CTx or can be made available through some backhaul links appears in [Devroye2006], [Goldsmith2009] and [Jovicic2009]. This asymmetric message knowledge at the two transmitters makes this channel very interesting and equally challenging as from the perspective of the PTx, this is an interference channel [Carleial1978] whereas it is similar to a broadcast channel as seen through the CTx. As is the case with the interference channel, the capacity region of this channel is unknown in general except for some special cases. The capacity region in the case of non-fading weak- interference was established in [Wu2007] and [Jovicic2009] whereas for the case of non-fading strong-interference, when both receivers can decode their respective interferences, the capacity region was determined in [Maric2007]. As overlay CR channel has great resemblance to the interference and the broadcast channels, most of the transmission schemes proposed for CR channels make use of the techniques proposed earlier for these aforementioned channels, the most notable of which are the rate splitting [Carleial1978] [Han1981] and the dirty paper coding (DPC) [Costa1983] [Weingarten2006].

We study the two user CR fading channel where the PTx and the PRx are the licensed spectrum holders and the cognitive pair, CTx and cognitive receiver (CRx), is the unlicensed pair and the CTx has the knowledge of the primary message. For the channel state information at the transmitter (CSIT), we assume that both transmitters know their channels toward PRx whereas both are ignorant of their channel realizations toward CRx. This assumption is motivated by the fact that the primary pair, being the license holder, is constantly communicating hence there are frequent transmissions from the PRx as well. This CSIT availability would clearly become straight-forward if the primary system is using time-division duplexing for its two-way transmissions. About the channel state information at the receiver (CSIR), both of the receivers are assumed to have the perfect knowledge about their channel realizations.

2.2.2 Assumptions and Contribution

We restrict the receiver strategies to single-user decoding which strikes out the use of rate-splitting [Carleial1978] from the pool of strategies. For the partial CSIT at the two transmitters and single stream transmission restriction, we propose two simple transmission schemes for this CR channel in “overlay paradigm”. The beamforming (BF) design at the multi-antenna CTx, based upon the channel state information (CSI) view that it has, is kept the same for both schemes. More specifically, the BF is done so that the PRx receives no interference from the cognitive message transmission. In the first scheme, the primary message is encoded independently at the two transmitters whereas in the second scheme, the CTx does the same primary encoding as carried out at the PTx. This allows coherent combining of the primary message at the PRx providing signal strength enhancement. Furthermore, this encoding makes the interference subspace at the CRx to lose its rank which makes it possible for the CRx to completely null it out with as few as two receive antennas. The achievable rate regions are characterized for both of the proposed schemes and performance comparison is provided with the rate region of an equivalent downlink (DL) channel which is obtained by combining both transmitters.

Notation: $\mathbb{E}$ denotes statistical expectation. Lowercase letters represent scalars, boldface lowercase letters represent vectors, and boldface uppercase letters denote matrices. $\mathbf{A}^\dagger$ denotes the Hermitian transpose of matrix $\mathbf{A}$. The identity matrix of n dimensions is denoted by $\mathbf{I_n}$. For an n-dimensional complex Gaussian vector $\mathbf{h}$, its norm $|\mathbf{h}|$ and squared-norm $|\mathbf{h}|^2$ are distributed as chi and chi-square with $2n$ degrees of freedom, denoted as χ_{2n} and χ_{2n}^2 respectively.

2.3 SYSTEM MODEL

The two user cognitive radio system is depicted in 2.1. The primary pair is equipped with single antenna each whereas the CTx and the CRx have n_t and n_r antennas respectively. The scalar and (row) vector channels at the PRx linking its antenna to PTx and CTx are denoted by h_{pp} and $\mathbf{h}_{\mathbf{pc}}^\dagger$ respectively. The (column) vector and matrix channels at the CRx linking its n_r antennas to PTx and CTx are denoted by $\mathbf{h_{cp}}$ and $\mathbf{H_{CC}}$ respectively. Every single channel entry is independent identically distributed (i.i.d.) standard complex Gaussian with zero mean and unit variance. The average power constraints of the PTx and the CTx are denoted by P_p and P_c respectively. If x_p and $\mathbf{x_c}$ denote the scalar and vector signals transmitted by PTx and CTx respectively, the scalar signal received at the PRx, denoted by y_p, is given by

$$y_p = \sqrt{P_p} h_{pp} x_p + \sqrt{P_c} \mathbf{h}_{\mathbf{pc}}^\dagger \mathbf{x_c} + z_p, \qquad \textbf{Eq 2.1}$$

where z_p denotes the scalar standard Gaussian noise at the PRx. Similarly the n_r-dimensional vector signal received at the CRx is given by the following equation

$$\mathbf{y_c} = \sqrt{P_p}\mathbf{h_{cp}}x_p + \sqrt{P_c}\mathbf{H_{cc}}\mathbf{x_c} + \mathbf{z_c}, \quad \textbf{Eq 2.2}$$

with each entry of the vector noise $\mathbf{z_c}$ at the CRx following standard Gaussian distribution. The PTx knows its message, denoted by $\mathcal{M}_\mathcal{P}$, and its channel to the PRx h_{pp}. The CTx knows its message $\mathcal{M}_\mathcal{C}$, the primary message $\mathcal{M}_\mathcal{P}$ and its channel to the PRx $\mathbf{h}_{\mathbf{pc}}^\dagger$. The CSIR is assumed to be perfect for both the PRx and the CRx.

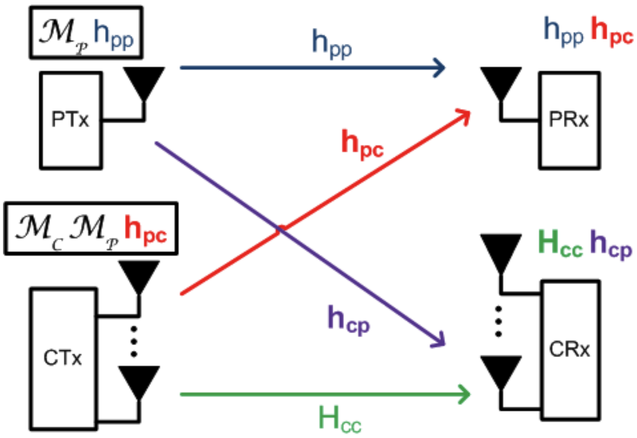

Figure 2.1. System model for 2-user cognitive radio interference channel.

2.4 MIMO BEAMFORMING (BF) DESIGN FOR OVERLAY PARADIGM

As the overlay paradigm of CR is the focus of this study, where the CTx has the knowledge of the primary message, it can help the primary pair achieve rates higher than that primary pair can achieve in the absence of the cognitive pair. The signal transmitted by the CTx $\mathbf{x_c}$ can be written as

$$\mathbf{x_c} = \mathbf{Wu} = \sqrt{\alpha}\mathbf{w_{cp}}u_{cp} + \sqrt{1-\alpha}\mathbf{w_{cc}}u_{cc}, \quad \textbf{Eq 2.3}$$

where $\mathbf{w_{cp}}$ and $\mathbf{w_{cc}}$ are the columns of the BF matrix $\mathbf{W}$ and denote the unit norm BF vectors at the CTx destined to PRx and CRx respectively. The primary and cognitive messages are encoded at the CTx so that u_{cp} and u_{cc} denote the standard Gaussian encoded signals (zero-mean unit-variance) intended for PRx and CRx and the parameter α controls the fractional power of the CTx dedicated for primary message. It is important to point out that the absence of full CSIT at the CTx makes it unable to predict the interference generated by $\mathcal{M}_\mathcal{P}$ at the CRx and hence DPC encoding [Costa1983] of the cognitive message against the primary message is not possible.

The only CSI available at the CTx is that of its channel to the PRx, $\mathbf{h_{pc}}$. This restricts very much the choice of beamforming design at the CTx and a very reasonable design for this CSI appears to choose $\mathbf{w_{cp}}$ as a matched filter (MF) beamformer to the PRx $\mathbf{h_{pc}}$ and to choose $\mathbf{w_{cc}}$ orthogonal to $\mathbf{h_{pc}}$ so that the PRx does not receive any interference from the message of the cognitive pair.

$$\mathbf{w_{cp}} = \frac{\mathbf{h_{pc}}}{\|\mathbf{h_{pc}}\|} \quad \textbf{Eq 2.4}$$

$$\mathbf{w_{cc}} =\perp \mathbf{h}_{\mathbf{pc}}^\dagger \Rightarrow \mathbf{h}_{\mathbf{pc}}^\dagger\mathbf{w_{cc}} = 0 \quad \textbf{Eq 2.5}$$

Although the design of $\mathbf{w_{cc}}$ seems to benefit fully the primary pair by producing no interference at the PRx, it does not particularly harm the signal at the CRx. The reason stems from the absence of the CSIT of CRx at the CTx which implies that, for any choice of the beamformer design, $\mathbf{w_{cc}}$ will stay independent of the direct channel $\mathbf{H_{cc}}$.

The PTx is a single antenna node hence it cannot apply any spatial filtering. So the signal transmitted by the PTx x_p is just the encoded primary message $\mathcal{M_P}$ in the form of standard Gaussian signalling u_{pp}.

$$x_p = u_{pp} \qquad \textbf{Eq 2.6}$$

2.5 INDEPENDENT PRIMARY MESSAGE ENCODING – ACHIEVABLE RATES

First we consider the case where the primary message $\mathcal{M_P}$ at the CTx is encoded independent of its encoding performed at the PTx. This implies that u_{pp} (encoded primary message at the PTx) and u_{cp} (encoded primary message at the CTx) are independent. The signal received at the PRx is given by

$$y_p = \sqrt{P_p} h_{pp} u_{pp} + \sqrt{\alpha P_c} \mathbf{h_{pc}^\dagger w_{cp}} u_{cp} + \sqrt{(1-\alpha)P_c} \mathbf{h_{pc}^\dagger w_{cc}} u_{cc} + z_p. \qquad \textbf{Eq 2.7}$$

The beamforming design at the CTx nulls out the interference signal at the PRx. So the received signal at the PRx becomes

$$y_p = \sqrt{P_p} h_{pp} u_{pp} + \sqrt{\alpha P_c} \|\mathbf{h_{pc}}\| u_{cp} + z_p. \qquad \textbf{Eq 2.8}$$

As the primary message has been encoded independently in u_{pp} and u_{cp}, the average achievable rate for the primary message $\mathcal{M_P}$ is given by

$$R_p = \mathbb{E}\log\left[1 + P_p |h_{pp}|^2 + \alpha P_c \|\mathbf{h_{pc}}\|^2\right], \qquad \textbf{Eq 2.9}$$

where the expectation needs to be taken over two channels.

As $|h_{pp}|^2 \sim \chi_2^2$ and $\|\mathbf{h_{pc}}\|^2 \sim \chi_{2n_t}^2$, the above rate becomes

$$R_p = \mathbb{E}\log\left[1 + P_p \chi_2^2 + \alpha P_c \chi_{2n_t}^2\right]. \qquad \textbf{Eq 2.10}$$

This rate expression clearly shows the advantage of the overlay paradigm as the primary pair can enjoy much higher rates as compared to what it can experience in the absence of the cognitive pair.

The vector signal received at the CRx is given by

$$\mathbf{y_c} = \sqrt{P_p} \mathbf{h_{cp}} u_{pp} + \sqrt{\alpha P_c} \mathbf{H_{cc} w_{cp}} u_{cp} + \sqrt{(1-\alpha)P_c} \mathbf{H_{cc} w_{cc}} u_{cc} + z_c. \qquad \textbf{Eq 2.11}$$

This equation shows that the CRx is receiving two interfering signals u_{pp} and u_{cp} and one desired signal u_{cc}. Thus the achievable rate of the cognitive pair would be highly dependent upon the number of antennas available at the CRx and its choice of receive filter. If we combine the interfering links as $\mathbf{H_i} = \left[\sqrt{P_p}\mathbf{h_{cp}} \quad \sqrt{\alpha P_c}\mathbf{H_{cc} w_{cp}}\right]$ and signals as $\mathbf{u_i} = [u_{pp} \quad u_{cp}]^{\mathrm{T}}$, the received signal at the CRx becomes

$$\mathbf{y_c} = \mathbf{H_i u_i} + \sqrt{(1-\alpha)P_c} \mathbf{H_{cc} w_{cc}} u_{cc} + \boldsymbol{z}_c. \qquad \textbf{Eq 2.12}$$

If the CRx is equipped with 3 or more antennas, it can null out its 2-dimensional interference $\mathbf{u_i}$. The projection matrix orthogonal to the CRx interference subspace, formed by the columns of $\mathbf{H_i}$, is $\mathbf{P} = \mathbf{I}_{\mathbf{n}_r} - \mathbf{H_i}\left(\mathbf{H_i^\dagger H_i}\right)^{-\mathbf{1}}\mathbf{H_i^\dagger}$. If the received signal $\mathbf{y_c}$ is projected onto this (orthogonal) subspace, it will zero-force (ZF) the interference streams. The rest of the receive dimensions can be used to MF the projected effective channel, given by $\mathbf{PH_{cc}w_{cc}}$. Hence the combined ZF-MF receiver which first zero-forces the interference and then does the MF to the orthogonally projected channel is given by

$$\mathbf{v}_\mathbf{c}^{\text{ZF-MF}} = \mathbf{w}_\mathbf{cc}^\dagger \mathbf{H}_\mathbf{cc}^\dagger \mathbf{P}^\dagger \mathbf{P} = \mathbf{w}_\mathbf{cc}^\dagger \mathbf{H}_\mathbf{cc}^\dagger \mathbf{P}, \qquad \textbf{Eq 2.13}$$

where the second equality is the result of the idempotent property of the projection matrices [Horn1985]. Due to well-known properties of the independent Gaussian vectors, it is known that the interference free filtered channel $\mathbf{v}_\mathbf{c}^{\text{ZF-MF}}\mathbf{H_{cc}w_{cc}}$ would be distributed as $\chi^2_{2(n_r-2)}$. This dictates that the achievable rate of the cognitive pair is given by:

$$R_c = \mathbb{E}\log\left[1 + (1-\alpha)P_c\chi^2_{2(n_r-2)}\right] \quad n_r \geq 3. \qquad \textbf{Eq 2.14}$$

If the CTx stops helping the primary pair ($\alpha = 0$), there is only one interfering stream and for $R_p = 0$, there is no interference. Thus the achievable rate at the CRx with ZF-MF are:

$$R_c = \begin{cases} \mathbb{E}\log\left[1 + (1-\alpha)P_c\chi^2_{2(n_r-2)}\right]; & n_r \geq 3, \alpha \neq 0 \\ \mathbb{E}\log\left[1 + P_c\chi^2_{2(n_r-1)}\right]; & n_r \geq 2, \alpha = 0, R_p \neq 0. \\ \mathbb{E}\log\left[1 + P_c\chi^2_{2n_r}\right]; & R_p = 0 \end{cases} \qquad \textbf{Eq 2.15}$$

Although ZF-MF receiver gives nice closed-form expressions for the rate of the cognitive pair, this is not necessarily the optimal receiver. The CRx can employ an SINR maximization filter, which we short-hand as MxSINR receiver, to get better rates. If CRx applies a unit norm receive filter $\mathbf{v_c}$ to the received signal $\mathbf{y_c}$ of (Eq. 2.12), the filtered signal becomes

$$\mathbf{v_c y_c} = \mathbf{v_c H_i u_i} + \sqrt{(1-\alpha)P_c}\,\mathbf{v_c H_{cc} w_{cc}} u_{cc} + \mathbf{v_c z_c}. \qquad \textbf{Eq 2.16}$$

The SINR of the desired signal at the CRx can be written as

$$\text{SINR}_\text{c} = \frac{\mathbf{v_c}\left[(1-\alpha)P_c \mathbf{H_{cc}w_{cc}w_{cc}^\dagger H_{cc}^\dagger}\right]\mathbf{v_c^\dagger}}{\mathbf{v_c}\left[\mathbf{H_i H_i^\dagger} + \mathbf{I}_{\mathbf{n}_r}\right]\mathbf{v_c^\dagger}}. \qquad \textbf{Eq 2.17}$$

The problem of finding the MxSINR receive filter which maximizes this SINR falls under the well-known paradigm of Generalized Eigen-Value problems (see [Sadek2007] and the references therein). Thus the MxSINR filter can be computed analytically as the maximum Eigen vector of the following matrix:

$$\left\{\left[\mathbf{H_i H_i^\dagger} + \mathbf{I}_{\mathbf{n}_r}\right]^{-\mathbf{1}}(1-\alpha)P_c \mathbf{H_{cc}w_{cc}w_{cc}^\dagger H_{cc}^\dagger}\right\}. \qquad \textbf{Eq 2.18}$$

This receiver will always exist contrary to ZF-MF receiver which requires more receive dimensions (antennas) than interference dimensions as its existence condition.

2.6 SAME PRIMARY MESSAGE ENCODING – ACHIEVABLE RATES

In this section, we treat the case when the encoding of the primary message $\mathcal{M}_\mathcal{P}$ is same at both the PTx and the CTx. Let this fact be taken into account by denoting both u_{pp} and u_{cp} as u_p. We further make a small change in the signal transmission at the PTx by nulling out the phase of the direct channel h_{pp} linking the PTx to the PRx. If $h_{pp} = |h_{pp}|e^{j\theta_{pp}}$, the signal transmitted at the PTx, x_p, is now given by

$$x_p = e^{-j\theta_{pp}}u_{pp} = e^{-j\theta_{pp}}u_p. \qquad \textbf{Eq 2.19}$$

This phase cancellation at the PTx does not involve any loss or benefit for the rate of the CRx but causes higher primary rates when the primary message coming from two transmitters is coherently combined at the PRx. Using the beamforming previously described, the signal received at the PRx is

$$y_p = \left(\sqrt{P_p}\left|h_{pp}\right| + \sqrt{\alpha P_c}\left\|\mathbf{h_{pc}}\right\|\right)u_p + z_p. \qquad \textbf{Eq 2.20}$$

And hence the achievable rate of the primary pair is given by

$$R_p = \mathbb{E}\log\left[1 + \left(\sqrt{P_p}\left|h_{pp}\right| + \sqrt{\alpha P_c}\left\|\mathbf{h_{pc}}\right\|\right)^2\right]. \qquad \textbf{Eq 2.21}$$

As $\left|h_{pp}\right| \sim \chi_2$ and $\left\|\mathbf{h_{pc}}\right\| \sim \chi_{2n_t}$, the above rate becomes

$$R_p = \mathbb{E}\log\left[1 + \left(\sqrt{P_p}\chi_2 + \sqrt{\alpha P_c}\chi_{2n_t}\right)^2\right]. \qquad \textbf{Eq 2.22}$$

A simple comparison of this achievable rate of the primary pair with the rate achieved without coherent combining, given in (Eq. 2.10), shows the promising gains of this scheme.

The received signal at the CRx is given by

$$\mathbf{y_c} = \left[\sqrt{P_p}\mathbf{h_{cp}}e^{-j\theta_{pp}} + \sqrt{\alpha P_c}\mathbf{H_{cc}}\mathbf{w_{cp}}\right]u_p + \sqrt{(1-\alpha)P_c}\mathbf{H_{cc}}\mathbf{w_{cc}}u_{cc} + \mathbf{z_c}. \qquad \textbf{Eq 2.23}$$

This equation shows the remarkable fact that the two interference streams coming from the two transmitters actually make a one dimensional subspace which makes it possible to null out the interference completely at the CRx even with two antennas. We again select the receive filter to be ZF-MF which first does ZF of the interference and then MF to the desired signal with the remaining dimensions. If we denote the interference vector as $\mathbf{h_i} = \left[\sqrt{P_p}\mathbf{h_{cp}}e^{-j\theta_{pp}} + \sqrt{\alpha P_c}\mathbf{H_{cc}}\mathbf{w_{cp}}\right]$, the orthogonal projection matrix is given by $\mathbf{P} = \mathbf{I}_{\mathbf{n}_r} - \mathbf{h_i}\left(\mathbf{h_i^\dagger}\mathbf{h_i}\right)^{-\mathbf{1}}\mathbf{h_i^\dagger}$ and the overall ZF-MF receive filter is

$$\mathbf{v_c^{\text{ZF-MF}}} = \mathbf{w_{cc}^\dagger}\mathbf{H_{cc}^\dagger}\mathbf{P}^\dagger\mathbf{P} = \mathbf{w_{cc}^\dagger}\mathbf{H_{cc}^\dagger}\mathbf{P}. \qquad \textbf{Eq 2.24}$$

The filtered signal at the CRx becomes

$$\mathbf{v_c^{\text{ZF-MF}}}\mathbf{y_c} = \sqrt{(1-\alpha)P_c}\mathbf{v_c^{\text{ZF-MF}}}\mathbf{H_{cc}}\mathbf{w_{cc}}u_{cc} + \mathbf{v_c^{\text{ZF-MF}}}\mathbf{z_c}, \qquad \textbf{Eq 2.25}$$

where the filtered noise $\mathbf{v_c^{\text{ZF-MF}}}\mathbf{z_c}$ is a standard Gaussian distributed scalar due to the independence of $\mathbf{v_c^{\text{ZF-MF}}}$ w.r.t. $\mathbf{z_c}$. The filtered channel $\mathbf{v_c^{\text{ZF-MF}}}\mathbf{H_{cc}}\mathbf{w_{cc}}$ is $\chi^2_{2(n_r-1)}$ distributed as only one direction $\mathbf{h_i}$ needs to be nulled out. It allows us to write the achievable rate of the cognitive pair as follows:

$$R_c = \begin{cases} \mathbb{E}\log\left[1 + (1-\alpha)P_c\chi^2_{2(n_r-1)}\right]; & n_r \geq 2, R_p \neq 0 \\ \mathbb{E}\log\left[1 + P_c\chi^2_{2n_r}\right]; & R_p = 0 \end{cases}. \qquad \textbf{Eq 2.26}$$

As we described for the previous strategy, a receive filter achieving higher rates than ZF-MF can be obtained as MxSINR receiver. This can be obtained by solving the Generalized Eigen-Value problem as detailed in the previous section. The MxSINR receiver would be the maximum Eigen vector of the following matrix:

$$\left\{\left[\mathbf{h_i}\mathbf{h_i^\dagger} + \mathbf{I}_{\mathbf{n_r}}\right]^{-\mathbf{1}}(1-\alpha)P_c\mathbf{H_{cc}}\mathbf{w_{cc}}\mathbf{w_{cc}^\dagger}\mathbf{H_{cc}^\dagger}\right\}. \qquad \textbf{Eq 2.27}$$

Note: If the PTx does not null out the phase as in (Eq. 2.19), the rate of the cognitive pair remains unchanged although the primary rate gets some hit as the effective channel to the PRx carrying primary message from two transmitters becomes $\mathbf{h_{pc}^\dagger}\mathbf{w_{cp}} + h_{pp} = \left\|\mathbf{h_{pc}}\right\| + h_{pp}$. The phase nulling operation carried out at the PTx makes the effective channel $\|\mathbf{h_{pc}}\| + h_{pp}$ which has higher mean, higher energy and hence results in larger achievable primary rate.

2.7 NUMERICAL AND SIMULATION RESULTS

In this section, we plot the rate regions which are achievable using the strategies proposed in the previous sections. The rate regions for various values of power constraints and when both CTx and CRx are equipped with 2 antennas each have been plotted in Figure 2.2. For comparison, we have also plotted the achievable rate regions for the equivalent DL channel which is obtained by collapsing both the PTx and the CTx into one node which has $(n_t + 1)$ antennas and its average power constraint is $P_p + P_c$, the sum of the constraints at the PTx and the CTx.

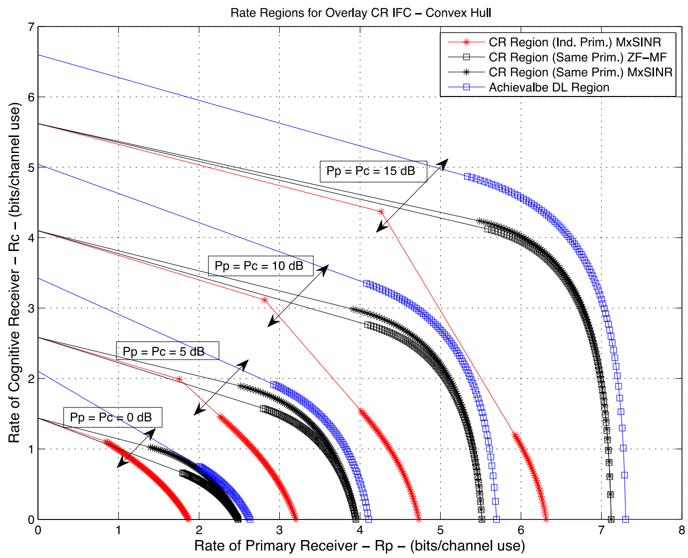

Figure 2.2. Both CTx and CRx have 2 antennas each. For same primary message encoding strategy, rates have been plotted when CRx is ZF-MF and MxSINR. With independent encoding, ZF is not possible so MxSINR CRx is used.

The CR regions for the same primary message encoding have been plotted for the two cases when the CRx employs ZF-MF and MxSINR receivers. We see that the rate regions corresponding to MxSINR receiver are larger but their gain over ZF-MF receiver regions diminishes at high SINRs. For independent primary message encoding scheme, the rate regions have only been plotted for MxSINR CRx as with 2 antenna CRx, perfect ZF of interference is not possible. At low power constraints, independent encoding scheme with MxSINR CRx gives some points which are outside of the region with same primary encoding followed by ZF-MF CRx. This is the consequence of the well-known inefficiency of ZF at low powers. We see that the same primary encoding scheme shows considerable rate advantage over independent encoding scheme which is due to the loss of coherent combining gain at the PRx in independent encoding. The rate regions of the same primary encoding strategy are close to the DL regions but there is significant difference in R_c for $R_p = 0$. This is the consequence of the relaxed DL power constraint as for $R_p = 0$, the cognitive message can be transmitted with power $P_p + P_c$ by the equivalent DL transmitter instead of true CTx power P_c. The achievable CR rate regions and DL rate regions have been plotted for $n_t = 3$, $n_r = 3$ in Figure 2.3 for various power constraints.

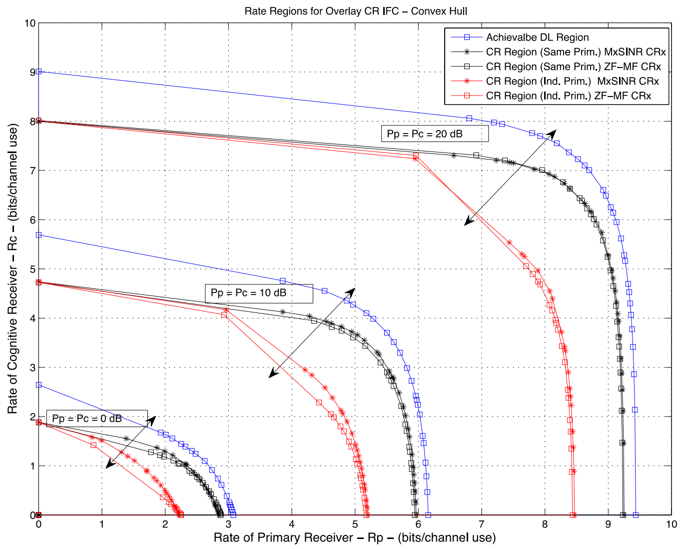

Figure 2.3. Both CTx and CRx have 3 antennas each. The rate regions have been plotted for both schemes when CRx employs Mx-SINR and ZF-MF receiver. MxSINR receivers bring some gains over ZF-MF for lower power constraints.

The CR rate regions for both schemes have been plotted for the two cases when the CRx is employing MxSINR receiver or ZF-MF receiver. As was the case in Figure 2.2, same primary message encoding strategy gives larger rate regions compared to independent encoding strategy.

2.8 CONCLUDING REMARKS

This section treats the Gaussian fading cognitive radio channel for single antenna primary pair and multi-antenna cognitive pair. The study is conducted with the availability of partial CSIT where each transmitter knows its channel to the PRx and has no information about its channel to the CRx. Two transmission strategies are proposed for this partial CSIT cognitive radio channel. The achievable rate regions are plotted and the performance comparison is provided with the achievable region of an equivalent DL channel which is obtained by collapsing both the PTx and the CTx into one. The strategy with same primary message encoding shows significantly larger rate regions over the strategy where the primary message is independently encoded at the two transmitters.

3 A NOVEL FRAMEWORK OF INTERFERENCE MITIGATION IN COGNITIVE NETWORKS

3.1 MAIN THEME

In this study we consider cognitive networks and focus on interference mitigation in the uplink. This section proposes an effective interference mitigation strategy based on centralized processing and exploitation of dominant interferences. The basic idea is two folds, i.e. using some linear processing to reduce the dimension of the cognitive system followed by some nonlinear processing to exploit dominant interferences. The resultant multistage receiver is characterized by the performance closer to that of an optimal receiver but with much less complexity.

3.2 INTRODUCTION

As wireless networks are inherently limited by their own interference, a lot of research focuses on interference reduction techniques, such as multiuser MIMO [Gesbert2007], interference alignment [Cadambe2008] or multi-cell processing [Gesbert2010]. Although these techniques can lead to considerable performance gains, it is unlikely that they will be able to carry the exponentially growing wireless data traffic. Due to this reason, either dense MIMO, i.e. network densification by increasing the number of antennas per unit area [Chandrasekhar2008] or massive MIMO, i.e. increasing the number of antennas at the base station (BS) [Marzetta2010] seems inevitable.

However the recent investigation that most of the voice and data communications take place indoors [Chandrasekhar2008] suggests against building huge infrastructures on the basis of economic viability for supporting these exploding rates over extended ranges. The viable solution appears to be network densification in the form of cell-size shrinking, i.e. femtocells [Chandrasekhar2008] which require little infrastructure expansion and can achieve manageable QoS and high data rates by allowing aggressive reuse of the spectrum. A femtocell is a small cellular area covering homes or offices, while a Femtocell Base Station (FBS) or Home eNodeB (HeNB) [3GPP TS 22.220] is a simple, low-cost miniature access point BS designed for indoor wireless service coverage of the corresponding macrocell.

Femtocell networks are considered as end-user deployed hotspots under the planned macrocell networks. With this, a two-tier network is formed, where the macro cell User Equipment (UE) are in tier-one, while the Femto User Equipments (FUE) are in tier-two. Since the two types of users are possibly using common frequency spectrum resources, inter-tier radio interference arises. The adhoc nature of femtocells further leads to intra-tier interference and both of these interferences act as a throughput constraining factor. This is even more crucial in dense urban scenarios where femtocells are massively deployed and cannot benefit from an efficient spatial reuse of the radio resources. Dealing with this interference demands extensive coordination at macro-level and sophisticated processing at FBS thereby overall increasing the network cost.

The main difference between a FBS and a Macro Base Station (Macro BS), as already introduced above, is that the former is user-deployed. That means that a FBS should be a “smart” device able to sense the environment and to adapt to its changes. This kind of “smart” devices has been defined by Mitola [Mitola2000] as Cognitive Radios (CRs) where the CRs are intelligent communication devices having the capability to co-exist with Primary license-holders.

The coexistence between a macrocell and a femtocell is depicted in Figure 3.1. Here the Cognitive User (CU), that denotes the FUE, wants to communicate with the Cognitive Receiver (CRx) that represents the FBS. This cognitive system coexists with a primary system where L Primary Users (PUs), i.e. the Macro UEs, want to talk to a common destination Primary Receiver (PRx), i.e. the Macro BS. The L PUs don’t care about the CU; hence they are producing a lot of interference at CRx (inter-tier interference).

The scenario in Figure 3.1 justifies the study of this section concerning interference management at the CRx side.

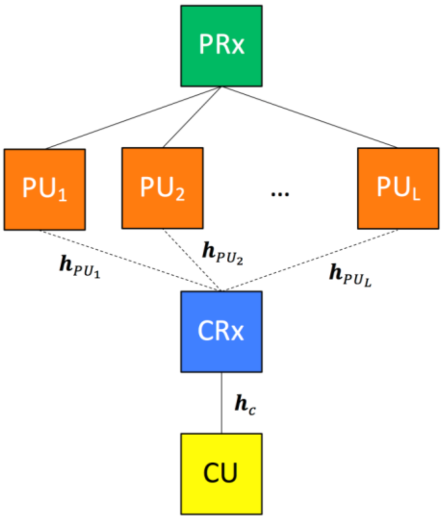

Figure 3.1. Coexistence between Cognitive and Primary systems.

3.2.1 Assumptions and Contribution

We focus on interference mitigation in the uplink of cognitive network based on Orthogonal Division Multiple Access (OFDMA). We consider an extension of the scenario depicted in Figure 3.1, where now we have more than one CRx, i.e. we have more than one femtocell. In this novel framework CRxs are connected to a CCS (Cognitive Control Station) via wired lines, such as fibers or coaxial cables, as shown in Figure 3.2.

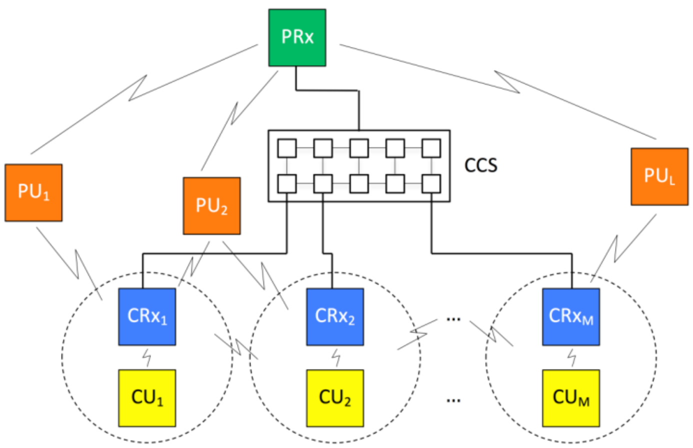

Figure 3.2. System Model: the uplink communication to CRx is interfered by neighboring CUs and Pus.

The working principle is to have each CRx as a simple terminal with analog amplifier which relays received signals to the CCS where the tasks of interference identification, correlation, and cancellation are performed. CCS will therefore have access to the signals of many CRxs thereby resulting into a virtual multi-antenna, multi-stream system and enabling centralized processing. However full blown up maximum likelihood (ML) centralized decoding at CCS will go beyond any acceptable limits of existing hardware. For this framework, we propose an interference mitigation strategy at CCS based on a multi-stage receiver which achieves the tradeoff of centralized processing (joint decoding) and complexity. The basic idea is amalgamation of linear and nonlinear processing in two stages, i.e. using some linear processing to reduce the dimension of the (cognitive) system in the first stage followed by some nonlinear processing to exploit dominant interferences in the second stage. This dual-stage processing is based on the classification of interferences into two groups, i.e. dominant interferences and weaker interferences and these two groups are processed in two different stages. Weaker interferences, being closer to Gaussianity [Annapureddy2009] compared to dominant interferences [Ghaffar2009], are attenuated in the first stage (linear receiver) which treats this group as Gaussian and reduces the system dimension. Stronger interferences are subsequently handled in the second stage (nonlinear receiver) which exploits the discrete structure of these dominant interferences. This two-stage receiver is characterized by low complexity due to the first stage being linear and is characterized by near-optimal performance due to the second stage being nonlinear and being based on the exploitation of interference structure. The proposed strategy therefore leads to the performance closer to that of an optimal receiver but with significant reduction in the complexity.

It is important to note that the receiver structure is fairly general and this dual stage receiver concept can be applied in any situation where the intended receiver is receiving interference from multiple sources which can be characterized into weaker and stronger interferers. Needless to say that this scenario is very important for cognitive radio receivers which would normally be operating over unlicensed bands and hence are prone to receiving multiple interfering signals.

Notation: $\mathbb{E}$ denotes statistical expectation. Lowercase and uppercase letters represent scalars, boldface lowercase letters represent vectors, and boldface uppercase letters denote matrices. $\mathbf{A}^{\mathrm{T}}$ $\mathbf{A}^{*}$ and $\mathbf{A}^{\dagger}$ denotes transpose, conjugate and conjugate transpose of matrix $\mathbf{A}$. $|\cdot|$ and $||\cdot||$ indicate norm of scalar and vector respectively. The identity matrix of n dimensions is denoted by$\mathbf{I_n}$.

3.3 SYSTEM MODEL

We consider the uplink of cognitive network as shown in Figure 3.2 where CRxs are connected with CCS via wired lines. Each CRx serves as an analog amplifier which amplifies and forwards the received signal to CCS where the PHY and MAC processing is carried out. CCS contains a series of interconnected Cognitive Boxes (CBs) with each CB connected to a CRx. In line with the on-going wireless standardizations as LTE [3GPP TS 36.211 Release 8] and LTE-Advanced [3GPP TS 36.211 Release 10], we consider that the system is based on OFDMA where the PUs employ BICM [Caire1998] based OFDM transmission. In the absence of any predefined spectrum management, the uplink cognitive communication is not only interfered by primary users but is also interfered by neighbouring cognitive transmissions. Assuming that the cyclic prefix of appropriate length is added to the OFDM symbols, the signal received at CB-1 at the *f*-th frequency tone after cascading IFFT at the CU and FFT at CB-1 is given as

$$\boldsymbol{y}_{1,f} = \boldsymbol{h}_{11,f}x_{1,f} + \boldsymbol{h}_{12,f}x_{2,f} + \cdots + \boldsymbol{h}_{1k,f}x_{k,f} + \boldsymbol{z}_{1,f} \qquad \textbf{Eq 3.1}$$

where $\boldsymbol{h}_{jl,f} \in \mathbb{C}^{n_r}$ is the vector characterizing Rayleigh fading channel from the *l*-th CU to the *j*-th CRx at the *f*-th frequency tone. Note that n_r denotes the number of receive antennas at the CRx.

$\boldsymbol{y}_{1,f}, \boldsymbol{z}_{1,f} \in \mathbb{C}^{n_r}$ are the vectors of the received symbols and circularly symmetric complex white Gaussian noise of double-sided power spectral density $N_0/2$ at n_r receive antennas of CRx-1. Interferences from relatively distant CUs being weaker have been merged in the noise [Annapureddy2009]. The complex symbol $x_{l,f}$ is transmitted by the *l*-th CU over a signal set $\chi_l \subseteq C$ with a Gray labeling map $\mu_l: \{0,1\}^{log_2|\chi_l|} \rightarrow \chi_l$ where $|\chi_l| = M_l$ is the cardinality of the constellation. Symbols from different CUs are assumed to be independent. Due to the interleaving operation in BICM, channels at different frequency tones are also assumed to be independent.

3.4 PROPOSED INTERFERENCE MITIGATION STRATEGY

At the CCS, an array of interconnected CBs is devised that centralizes the signal processing. The interference for one CB is the desired signal for another CB so some common processing or the exchange of information between the CBs can help in mitigating the interference. In this context, multi-user detection strategy can be adopted at the CCS through ML receivers but though being optimal, it will lead to enormous complexity. On the other hand, single-user detection strategy can be adopted through single-user receivers which would be less complex but will be highly suboptimal. Need is to devise an effective interference mitigation strategy which can be implemented within the limits of existing hardware complexity constraints. To this end, we propose a multi-stage receiver in each CB which on one hand reaps the benefits of multi-user detection but on the other hand is not as complex as multi-user detectors. Based on the idea that CCS can process the received signals jointly, we propose that the received signals at different CBs are categorized into different interference groups and then these groups are handled by multi-stage receivers.

Without loss of generality, we consider CB-1 whose desired signal is x_1 but it is encountering dominant interferences as $x_2, \ldots, x_k$ which are either from the adjacent CUs or PUs. The complexity of ML detection at CB-1 for detecting the desired signal would be $\mathcal{O}(|\chi|^k)$. Disregarding the subcarrier index and merging the weaker interferences in noise, the signal at CB-1 is given as

$$\boldsymbol{y}_1 = \boldsymbol{h}_{11}x_1 + \cdots + \boldsymbol{h}_{1k}x_k + \boldsymbol{z}_1. \qquad \textbf{Eq 3.2}$$

To detect x_1, CB-1 accesses the signals from other CBs which have x_1 as dominant interferer. Let the signal of *l*-th neighbouring CRx is given as

$$\boldsymbol{y}_l = \boldsymbol{h}_{l1}x_1 + \cdots + \boldsymbol{h}_{ln}x_n + \boldsymbol{z}_l. \qquad \textbf{Eq 3.3}$$

CB-1 stacks these signals in a column vector which is given as

$$\begin{bmatrix} \boldsymbol{y}_1 \\ \vdots \\ \boldsymbol{y}_m \end{bmatrix} = \begin{bmatrix} \boldsymbol{h}_{11} & \ldots & \boldsymbol{h}_{1k} & 0 & \cdots & 0 \\ \vdots & \vdots & \vdots & \vdots & \vdots & \vdots \\ \boldsymbol{h}_{m1} & \ldots & \ldots & \ldots & \cdots & \boldsymbol{h}_{mn} \end{bmatrix} \begin{bmatrix} x_1 \\ \vdots \\ x_k \\ \vdots \\ x_n \end{bmatrix} \qquad \textbf{Eq 3.4}$$

which can be written as

$$\boldsymbol{y} = \boldsymbol{h}_1x_1 + \cdots + \boldsymbol{h}_kx_k + \cdots + \boldsymbol{h}_nx_n + \boldsymbol{z} = \boldsymbol{Hx} + \boldsymbol{z}. \qquad \textbf{Eq 3.5}$$

Note that the first column of $\boldsymbol{H}$ corresponding to x_1 will have all non-zero entries while other columns will have some entries as zero. This is due to the fact that all signals are not received by all CRxs and the received signals which have been grouped at CB-1 are based on the criteria that they all have x_1 as a dominant signal. The complexity of ML detection of x_1 in (Eq. 3.5) is $\mathcal{O}(|\chi|^n)$ where $n \geq k$. Though additional spatial diversity has been obtained by combining signals, the complexity of detection has also increased fundamentally. Based on the sparsity of $\boldsymbol{H}$, we propose a multi-stage receiver for the detection of x_1.

3.5 MULTI-STAGE RECEIVER

There are several choices of receivers in such an interference-limited system which can be broadly divided into two groups. First group of receivers are linear receivers which are based on the simplifying Gaussian assumption for the alphabets. Design criteria of these receivers is largely influenced by the practical constraints where complexity is the limiting factor in receive processing. These constraints have led to naive linear solutions, which are based on neglecting the integer constraint coupled with the simplifying Gaussian assumption for the desired signal and interfering signals and then subsequently projecting the so-obtained solution onto the finite set of permissible alphabets of desired signal. These linear receivers as MMSE [Ghaffar2009] and zero forcing (ZF) [McKay2007] in general work poorly especially in the case of limited number of alphabets when central limit theorem dictating the overall sum of alphabets to be Gaussian does not hold. Second group of receivers are non-linear receivers as optimal ML receivers or near optimal sphere decoders [Brunel2004] or lattice reduction aided receivers [Yao2002] etc. These receivers are characterized by high complexity though their performance is much better than their linear counterparts. These receivers consider the alphabets to be from the realistic finite constellations and exploit their discrete structure in the detection process. This adds significantly to the detection complexity which is exponential in the number of alphabets, i.e. dimension of the system. Due to this complexity, these receivers are generally not preferred in practical systems [Larsson2008], rather they serve as an upper bound on the system performance.

We here propose a split detection scheme which comprises of two stages. This splitting is based on different behavior of interferences as per their strength. Weaker interferences are closer to Gaussianity and exploiting the structure of these interferences will only lead to limited gains but at the cost of huge increase in complexity. Ignoring these interferences and merging them in noise, i.e. single-user detection, will lead to considerable degradation in the performance. These interferences can however be effectively attenuated by linear processing. On the other hand, exploiting the structure of dominant interferences lead to significant improvement in the performance. Though the complexity increases, but the performance gains justify this additional complexity. We therefore propose a multi-stage receiver with the first stage being linear where relatively weaker interferences are mitigated, assuming them to be Gaussian and the system dimension is adapted so as to be feasible for the subsequent stage. The second stage is a non-linear ML receiver which exploits the structure of dominant interferences in the detection of the desired stream.

3.5.1 Inner Receiver

In the proposed multi-stage receiver, inner receiver is designed to reduce the dimension of the system suitable for the outer ML receiver. It is devised on the basis of MMSE criteria [Tse2005] with the task of mitigating relatively weaker interferences and therefore adapting the dimension of the system suitable for the outer ML receiver. Amongst the dominant interferences, a group of relatively weaker interferences is categorized and the inner receiver attenuates this group. This classification is based on the norms of channel vectors so the interfering channels incorporating more zero entries are segregated as weaker interferences. Without loss of generality, let's suppose that $x_{k+1}, \dots, x_n$ streams are categorized as weaker interferences. So a filter based on MMSE criteria to suppress these weaker interference group would be given as

$$\boldsymbol{W} = \boldsymbol{R}_{y,y}^{-1}\boldsymbol{R}_{y,\underline{\boldsymbol{H}}\,\underline{\boldsymbol{x}}} = \left(\underline{\boldsymbol{H}}\,\underline{\boldsymbol{H}}^{\dagger} + \overline{\boldsymbol{H}}\,\overline{\boldsymbol{H}}^{\dagger} + N_0\boldsymbol{I}_{n_r}\right)^{-1}\left(\underline{\boldsymbol{H}}\,\underline{\boldsymbol{H}}^{\dagger}\right) \qquad \textbf{Eq 3.6}$$

where $\boldsymbol{R}_{y,y}$ is the covariance matrix of $\boldsymbol{y}$, $\underline{\boldsymbol{H}} = [\boldsymbol{h}_1 \quad \dots \quad \boldsymbol{h}_k]$ and $\overline{\boldsymbol{H}} = [\boldsymbol{h}_{k+1} \quad \dots \quad \boldsymbol{h}_n]$. Application of the MMSE filter and the Gaussian assumption of the post detection interference would lead to

$$\boldsymbol{W}^{\dagger}\boldsymbol{y} = \boldsymbol{W}^{\dagger}\underline{\boldsymbol{H}}\,\underline{\boldsymbol{x}} + \boldsymbol{W}^{\dagger}\overline{\boldsymbol{H}}\,\overline{\boldsymbol{x}} + \boldsymbol{W}^{\dagger}\boldsymbol{z} = \boldsymbol{W}^{\dagger}\underline{\boldsymbol{H}}\,\underline{\boldsymbol{x}} + \boldsymbol{z}' \qquad \textbf{Eq 3.7}$$

where $\underline{\mathrm{x}} = [\mathrm{x}_1 \quad \cdots \quad \mathrm{x}_k]^{\mathrm{T}}$ and $\overline{\mathrm{x}} = [\mathrm{x}_{k+1} \quad \cdots \quad \mathrm{x}_n]^{\mathrm{T}}$. The variance of z', i.e. interference plus noise is given by $N_1 = \mathbb{E}(\|\boldsymbol{z}'\|^2 \mid \boldsymbol{H})$. Knowing that $\overline{\boldsymbol{H}}\,\overline{\boldsymbol{H}}^{\dagger} = \boldsymbol{H}\,\boldsymbol{H}^{\dagger} - \underline{\boldsymbol{H}}\,\underline{\boldsymbol{H}}^{\dagger}$, the covariance matrix of $\boldsymbol{z}'$ is given as

$$\begin{aligned}\boldsymbol{\Psi} &= \mathbb{E}\left[\left(\boldsymbol{W}^{\dagger}\overline{\boldsymbol{H}}\,\overline{\boldsymbol{x}} + \boldsymbol{W}^{\dagger}\boldsymbol{z}\right)\left(\boldsymbol{W}^{\dagger}\overline{\boldsymbol{H}}\,\overline{\boldsymbol{x}} + \boldsymbol{W}^{\dagger}\boldsymbol{z}\right)^{\dagger}\right] = \\ &= \underline{\boldsymbol{H}}\,\underline{\boldsymbol{H}}^{\dagger}\boldsymbol{A}^{-1}\underline{\boldsymbol{H}}\,\underline{\boldsymbol{H}}^{\dagger} - \underline{\boldsymbol{H}}\,\underline{\boldsymbol{H}}^{\dagger}\boldsymbol{A}^{-1}\underline{\boldsymbol{H}}\,\underline{\boldsymbol{H}}^{\dagger}\boldsymbol{A}^{-1}\underline{\boldsymbol{H}}\,\underline{\boldsymbol{H}}^{\dagger}\end{aligned} \qquad \textbf{Eq 3.8}$$

where $\boldsymbol{A} = \left(\underline{\boldsymbol{H}}\,\underline{\boldsymbol{H}}^{\dagger} + \overline{\boldsymbol{H}}\,\overline{\boldsymbol{H}}^{\dagger} + N_0\boldsymbol{I}_{n_r}\right)$.

3.5.2 Outer Receiver – ML

After the adaptation of system dimension by the inner receiver, outer receiver focuses on the desired stream x_1. Here we propose the practical version of ML receiver, i.e. max log MAP receiver [Caire1998] as the outer receiver. Detailed discussion of this receiver will go out of the scope of this deliverable, please consult [Caire1998] for details. Bit metric for the bit *b* at the *i*-th location of the first (desired) spatial stream at CB-1 is given as

$$\begin{aligned}\lambda_1^i(\boldsymbol{W}^{\dagger}\boldsymbol{y}, b) &\approx \min_{\substack{x_1 \in \chi_{1,b}^i \\ x_2 \in \chi_2, \ldots, x_k \in \chi_k}} \left[\frac{1}{N_1}\left\|\boldsymbol{W}^{\dagger}\boldsymbol{y} - \boldsymbol{W}^{\dagger}\boldsymbol{h}_1 x_1 - \cdots - \boldsymbol{W}^{\dagger}\boldsymbol{h}_k x_k\right\|^2\right] \\ &= \min_{\underline{\boldsymbol{x}} \in \chi_b^i}\left[\frac{1}{N_1}\left\|\boldsymbol{W}^{\dagger}\boldsymbol{y} - \boldsymbol{W}^{\dagger}\underline{\boldsymbol{H}}\,\underline{\boldsymbol{x}}\right\|^2\right]\end{aligned} \qquad \textbf{Eq 3.9}$$

where $\chi_{1,b}^i$ denotes the subset of the signal set $x_1 \in \chi_1$ whose labels have the value $b \in \{0,1\}$ in the position *i*. Note that $N_1 = tr(\boldsymbol{\Psi})$ where $tr[\cdot]$ refers to the trace of matrix. Bit LLRs are calculated using these bit metrics which are then deinterleaved and are subsequently fed to the decoders for the detection of the desired stream. The detection complexity of this ML receiver is of the order of *k*, i.e. $\mathcal{O}\left(|\chi|^k\right)$.

3.5.3 MMSE based detection

For comparison purposes, we consider the state of the art parallel detection scheme in such scenarios, i.e. linear MMSE receiver [Tse2005] which is given as

$$\boldsymbol{M} = \left(\boldsymbol{H}\boldsymbol{H}^{\dagger} + N_0\boldsymbol{I}_{n_r}\right)^{-1}\boldsymbol{H} \qquad \textbf{Eq 3.10}$$

whose application yields

$$\boldsymbol{M}^{\dagger}\boldsymbol{y} = \boldsymbol{H}^{\dagger}\boldsymbol{A}^{-1}\boldsymbol{H}\boldsymbol{x} + \boldsymbol{H}^{\dagger}\boldsymbol{A}^{-1}\boldsymbol{z}. \qquad \textbf{Eq 3.11}$$

The *l*-th component is given as

$$\boldsymbol{h}_l^{\dagger}\boldsymbol{A}^{-1}\boldsymbol{y} = \boldsymbol{h}_l^{\dagger}\boldsymbol{A}^{-1}\boldsymbol{h}_l x_l + z' \qquad \textbf{Eq 3.12}$$

The variance of z', i.e. interference plus noise is given by $N_l = \mathbb{E}(|z'|^2 \mid \boldsymbol{H})$. Assuming that $\boldsymbol{U}\boldsymbol{U}^{\dagger} = \boldsymbol{H}\boldsymbol{H}^{\dagger} - \boldsymbol{h}_l\boldsymbol{h}_l^{\dagger}$, we get

$$N_l = \boldsymbol{h}_l^{\dagger}\boldsymbol{A}^{-1}\boldsymbol{U}\boldsymbol{U}^{\dagger}\boldsymbol{A}^{-1}\boldsymbol{h}_l + N_0\boldsymbol{h}_l^{\dagger}\boldsymbol{A}^{-1}\boldsymbol{A}^{-1}\boldsymbol{h}_l = \boldsymbol{h}_l^{\dagger}\boldsymbol{A}^{-1}\boldsymbol{h}_l - \left|\boldsymbol{h}_l^{\dagger}\boldsymbol{A}^{-1}\boldsymbol{h}_l\right|^2. \qquad \textbf{Eq 3.13}$$

So the SINR is given as

$$\mathrm{SINR} = \frac{\left|\boldsymbol{h}_l^{\dagger}\boldsymbol{A}^{-1}\boldsymbol{h}_l\right|^2}{\boldsymbol{h}_l^{\dagger}\boldsymbol{A}^{-1}\boldsymbol{h}_l - \left|\boldsymbol{h}_l^{\dagger}\boldsymbol{A}^{-1}\boldsymbol{h}_l\right|^2} = \frac{\left|\boldsymbol{h}_l^{\dagger}\boldsymbol{A}^{-1}\boldsymbol{h}_l\right|}{\boldsymbol{1} - \left|\boldsymbol{h}_l^{\dagger}\boldsymbol{A}^{-1}\boldsymbol{h}_l\right|}. \qquad \textbf{Eq 3.14}$$

Note that $\boldsymbol{A}^{-1}$ being a positive definite matrix, $\boldsymbol{h}_l^{\dagger}\boldsymbol{A}^{-1}\boldsymbol{h}_l$ is a positive real number. The bit metric for the bit *b* at the *i*-th location of the *l*-th spatial stream is given as

$$\lambda_1^i(\boldsymbol{M}^{\dagger}\boldsymbol{y}, b) \approx \min_{x_l \in \chi_{l,b}^i}\left[\frac{1}{N_l}\left|\boldsymbol{h}_l^{\dagger}\boldsymbol{A}^{-1}\boldsymbol{y} - \boldsymbol{h}_l^{\dagger}\boldsymbol{A}^{-1}\boldsymbol{h}_l x_l\right|^2\right] \qquad \textbf{Eq 3.15}$$

3.6 NUMERICAL AND SIMULATION RESULTS

For simulations, we consider the uplink of a cognitive network where transmission is based on BICM OFDM. We consider single antenna CUs and dual-antenna CRxs. The CUs employ rate-$1/3$ LTE turbo code [3GPP TS 36.212] punctured to rate $1/2$. We consider ideal OFDM system (no ISI) and analyse the system in the frequency domain where the channel has i.i.d. Gaussian matrix entries with unit variance. We consider fast fading where the channel is independently generated for each channel use. We focus on the detection of the signal of CU_1 at CB-1, i.e. x_1 and we look at the frame error rate (FER) where the frame length is fixed to 1056 information bits. The signal x_1 is directly interfered by 2 signals from neighboring CUs. CB-1 at CCS stacks the signals from other CBs as per (Eq 3.4) and (Eq 3.5) where x_1 is the dominant signal.

This leads to $n = 6$, i.e. the system dimension increases to 6 (one desired signal x_1 and five interfering signals). For reception, we consider the proposed multi-stage receiver and vary the distribution of streams into two groups. Once all 5 interferences are included in the first group, then multi-stage receiver breaks down to the standard linear MMSE receiver whereas when all 5 interferences are included in the second group, the multi-stage receiver transforms to the full blown-up ML receiver.

Figure 3.3 shows the results for different distributions of interferences in two groups. Note that interferences are ordered in increasing strength at the multi-stage receiver and stronger interferences are the ones which are firstly shifted to the second group. The results show that there is a progressive improvement in the performance as the size of second group increases which is obvious as the multi-stage receiver gets closer to the ML receiver with the increase of the size of second group, however the complexity of detection also increases. The extent of improvement in the performance decreases as more interferences are put into the second group. This is due to the fact that exploiting the structure of weaker interferences leads to only marginal performance improvements. Note that there is an improvement of about 4 dB when 2 dominant interferences are put into second group where the complexity of detection increases from $\mathcal{O}(|\chi|)$ to $\mathcal{O}(|\chi|^3)$. Putting more interferences (weaker) in the second group give a gain of merely 1 dB whereas the complexity of detection increases to $\mathcal{O}(|\chi|^6)$. These results show that there is an appropriate tradeoff between the performance and complexity of multi-stage receiver and the performance gets quite close to the optimal ML receiver with much reduced complexity. For comparison purposes, we have also included the results for single-user receiver when the detection is carried out at CRx_1. This comparison shows the significant gains that can be achieved by centralized processing at the CCS as compared to stand-alone processing at CRxs.

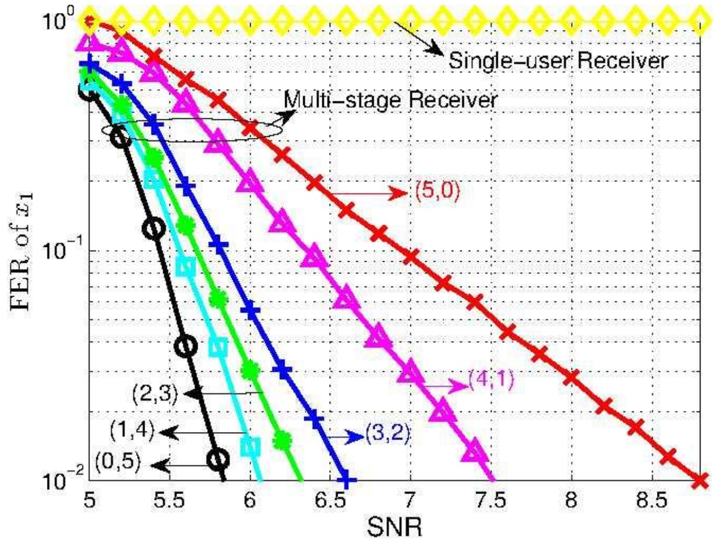

Figure 3.3. Multi-stage receiver at CCS to detect $\mathbf{x_1}$. $(\mathbf{a}, \mathbf{b})$ indicates splitting of interferences into groups where a indicates the number of interferences in the first group (linear receiver) while b indicates the number of interferences in the second group (non-linear ML receiver). Alphabets belong to QPSK and SNR is the received SNR at CRx1.

3.7 CONCLUDING REMARKS

This section focuses on interference mitigation in a cognitive network based on dual stage reception strategy. We have proposed the categorization of interferences into weaker and stronger groups. In the proposed multistage receiver, the first stage is linear which reduces the system dimension by attenuating weaker interferences. After the adaptation of system dimension, the subsequent stage is a non-linear ML receiver which exploits the structure of dominant interferences in the detection process. This proposed strategy achieves near ML performance with much reduced complexity.

4 CONCLUSIONS

This deliverable documents the MIMO transmission and reception algorithms particularly suited for cognitive radio systems. A special attention is paid to the overlay regime of cognitive radios which is the most challenging and consequently most rewarding of the cognitive regimes.

First we presented a scheme for a simple case of two user Gaussian fading Cognitive Radio (CR) interference channel. It has two pairs of communication terminals, one representing primary license holder and the other as unlicensed cognitive pair. The cognitive pair is equipped with multiple antennas. The channel state information is assumed partial such that each transmitter knows its channel to the PRx but has no information about its channel to the CRx. Two simple transmission strategies are proposed focusing the CR channels in the so-called “overlay paradigm” where the CTx not only transmits its own message but helps as well transmitting the primary message. The simulations results are presented in terms of spectral efficiency of the cognitive radio systems with the proposed schemes.

In the second section, the focus is turned to interference mitigation in the uplink of cognitive networks based on centralized processing. Due to the presence of primary system and possibility of multiple cognitive systems, this scenario is of high concern. We have proposed the categorization of interferences into weaker and stronger groups. We have further proposed a multistage receiver where the first stage is linear which reduces the system dimension by attenuating weaker interferences. After the adaptation of system dimension, the subsequent stage is a non-linear ML receiver which exploits the structure of dominant interferences in the detection process. The simulation results with the proposed strategy show near ML performance with much reduced complexity making the proposed dual stage receiver a suitable candidate for implementation into practical systems.

5 ACRONYMS

Term	**Description**
3GPP	3rd Generation Partnership Project
BF	Beamforming
BICM	Bit-Interleaved Coded Modulation
BS	Base station
CB	Cognitive Box
CCS	Cognitive Control Station
CR	Cognitive Radio
CRx	Cognitive Receiver
CSIR	Channel State Information at the Receiver
CSIT	Channel State Information at the Transmitter
CTx	Cognitive Transmitter
CU	Cognitive User
DL	Downlink
DPC	Dirty Paper Coding
FBS	Femtocell Base Station
FC	Femtocell
FER	Frame Error Rate
FFT	Fast Fourier Transform
FUE	Femto User Equipment
HeNB	Home eNodeB
IFC	Interference Channel
IFFT	Inverse Fast Fourier Transform
ISI	Inter Symbol Interference
LTE	Long Term Evolution
LTE-A	Long Term Evolution – Advanced
MF	Matched Filter
MIMO	Multiple Inputs Multiple Outputs
ML	Maximum Likelihood
MMSE	Minimum Mean Square Error

Term	**Description**
OFDM	Orthogonal Frequency Division Multiplexing
OFDMA	Orthogonal Frequency Division Multiple Access
PRx	Primary Receiver
PTx	Primary Transmitter
PU	Primary User
QoS	Quality Of Service
SINR	Signal to Interference plus Noise Ratio
TVWS	TV White Spaces
UE	User Equipment
WP	Work Package
ZF	Zero Forcing Filter

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